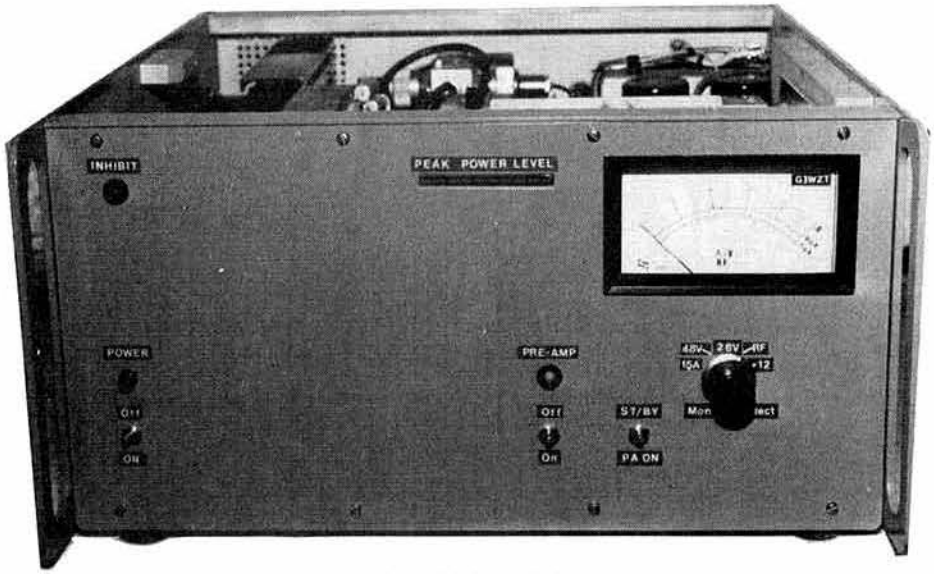
GUIDELINES



FOR THE DESIGN OF SEMICONDUCTOR VHF POWER AMPLIFIERS

BY JOHN MATTHEWS, G3WZT

My intention in writing this article is to provide an introduction to some of the techniques which anyone can employ in the design of high power solid-state vhf linear amplifiers. I have included a tried and tested means of power dividing and combining for narrow band applications, and simple methods of input/output matching using quarter wave coaxial transmission line transformers.

All of the examples are based upon a working 144MHz 250W pep power amplifier stage and a 75W pep driver with integral power supplies. Description of the driver stage will show a simple means of providing the required push-pull anti-phase signals using quarter wave coaxial matching transformers.

Detailed component and constructional details are not given as the devices used in the pa stages are now classed as obsolete and could prove to be both difficult and expensive to obtain. Using the various methods to be described it will, however, be possible for an experienced constructor to design amplifiers from first principles using transistors which can be found at sensible prices!

One unfortunate aspect of high power solid state amplifiers is their inability to withstand high values of vswr at normal operating power levels. This depends to a large extent on the magnitude of the mismatch, but 'firing up' at full drive power and no load can result in instant destruction and a hefty repair bill!

The design described here includes vswr protection which may be set to a pre-determined level. When this level is exceeded, the amplifier is disabled and the otherwise inevitable catastrophe avoided.

WHY TRANSISTORS?

I am sure that most present owners of 250W amplifiers will have already asked themselves whether it's wise to take the semiconductor route. The devices used in this pa were obtained some time ago and had been taking pride of place in my 'goodies box'. It just seemed a great shame not to use them; I also assumed that I would learn something along the way about the design of high power, vhf solid state amplifiers.

At this power level, the 4CX250 is undoubtedly the most popular and economical provider of rf energy at 144MHz. This design is an alternative approach requiring a regulated supply of only 48V. Without doubt, the solid state approach is more complex and expensive than its valve counterpart but if properly designed and driven, is capable of providing very good performance.

Unfortunately, solid state amplifiers have earned themselves a very poor reputation – a problem which appears to stem from consistent overstatement of performance by most, if not all, commercial manufacturers of solid state linear power amplifiers. In the past few years I have had the opportunity to performance test several makes and models of commercially manufactured linear amplifiers, and found very few capable of producing their quoted output power. Judging by many signals heard on 144MHz, some operators seem to be able physically to achieve the theoretical third order intercept point of a power amplifier! Careful reading of the small print of published data sheets shows quoted output powers of ±0·5dB and others ±1dB. Do not be surprised if your new amplifier rated at 175W only delivers 140W – at ±1dB it's still inside the specification. However hard you drive it makes no difference, the output does not get bigger – only wider! The problem is compounded by a total lack of specification for linearity. If the quoted output power can be met and increasing the drive power further makes no improvement in output, the amplifier is without doubt saturated and not suitable for linear service at this power level.

Drive your pa with sufficient power to give 10-15% below maximum available output – but *never* more than that, not even on speech peaks – and you should keep everyone happy. Bar graph displays are a valuable asset as they do not suffer ballistic problems of moving coil meters when trying to monitor peak output power.

POWER OUTPUT STAGE

The output stage uses two BLX15 bipolar transistors operating from a 48V supply. Although these devices are early generation types, they still provide creditable performance and high rf output at 144MHz. In common with many older vhf high power bipolar devices the gain is modest. At the frequency in question, 5–7dB power gain is all that can be expected. This places quite high demands on the driver stage which must be capable of delivering 60–70W pep with good linearity.

Various configurations for the output stage were considered and the decision made to use two identical stages in parallel with 3dB hybrid directional couplers to divide and combine power from each stage. A simple block diagram of the amplifiers and 3dB hybrids is shown in Fig 1.

INPUT MATCHING

Manufacturers data for the BLX 15 shows the input impedance to be 0·3+j1·80ohms. These are series components and quite typical of high power bipolar devices. At first sight this would seem to be a rather difficult impedance to match into a 50ohm system, but, if these values are converted to their equivalent parallel values, the problem looks far more manageable.

Converting into parallel values (see reference 1) yields 11 ohms resistive and 1·85ohms inductive reactance. This equates to an 11ohm resistor in parallel with a 2nH inductor at 145MHz. In order to make this complex impedance look purely resistive, a component of the opposite sign must be placed in parallel. Capacitive reactance of 1·85ohms at 145MHz is 593pF and is the theoretical value required to make the BLX15 input look purely resistive (Fig 2).

In practice, it was found that a slightly smaller value was required for best match. This is due to small variations in manufacturers' quoted values for input impedance and the practical realisation of calculated values in the T match network. A total of 440pF is used, part of which is made variable to optimise the input match. All that needs to be done to complete the pa input network is to provide a means of matching the 11ohm input resistance to the 50ohms required by the power divider. This is accomplished using a quarter wave coaxial transmission line transformer.

QUARTER WAVE COAXIAL LINE TRANSFORMERS

There are many clever tricks which can be used with coaxial cables to perform impedance transformation and matching, but for the purpose of this article only simple quarter wave types will be used. The short explanation that follows is based upon the needs of the amplifier described but is applicable to other frequencies and transformation ratios.

Source, load, and line impedance, are related in the following way.

$$Zline = \sqrt{Zsource \times Z\,load} \quad \ldots 1$$

Where Zline is the unknown, required coaxial line impedance. Zsource is 50ohms required at the power divider ports and, Zload is 11ohms, represented by the transistor input.

$$\text{Therefore } Zline = \sqrt{50 \times 11} = 23{\cdot}45\Omega \quad \ldots 2$$

So where do we get 23·4ohms coaxial cable? The simple answer is nowhere; but if two lengths of 50ohms coax are paralleled the end result is 25ohms which is quite close enough. Rearranging equation (1) around for Zsource gives:

$$Zsource = (Zline)2 / Zload \quad \ldots 3$$

This means that the impedance seen at the input to the matching section when connected to the transistor will be:

$$Zsource = 25^2 / 11 = 56{\cdot}8\Omega \quad \ldots 4$$

This assumes that the BLX15 input is exactly 11ohms and the coax chosen is exactly 25ohms, and one quarter wavelength long. The resulting errors are negligible and from a practical viewpoint can be ignored.

LINE LENGTH

In order to meet the requirements of equation 1, the transformer line length must be one quarter wavelength. It should be remembered that line length required is *electrical length*, and the free space value must be multiplied by the velocity factor of cable used.

For ptfe dielectric the value is 0·70, and for polyethylene (pe), 0·66. For example, using ptfe dielectric the required 25ohms transformer line length will be:

$$300/FMHz \times 0{\cdot}70/4 = 300/145 \times 0{\cdot}172 = 35{\cdot}6cm \quad \ldots 5$$

A polar plot for this type of transformer is shown in Fig 3. One end of the transformer was terminated with 12·5ohms and the plot shows changing characteristics of the transformer between 50MHz and 450MHz. 145MHz is located close to the centre, indicating a good match into a 50ohm system. The inner circle on the polar plot shows a constant vswr circle of 1·5:1. From this, the upper and lower frequencies corresponding to a 1·5:1 vswr are seen to be 160 and 114MHz respectively. This gives a graphical representation of typical bandwidths which may be expected from this type of transformer.

Reference 2 is recommended if further understanding of transmission line transformers is required.

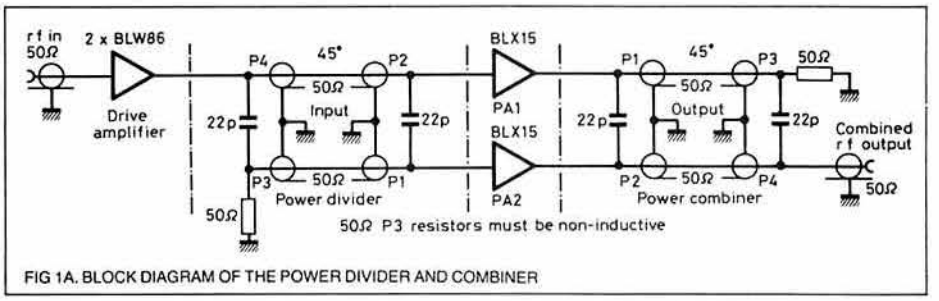


FIG 1A. BLOCK DIAGRAM OF THE POWER DIVIDER AND COMBINER

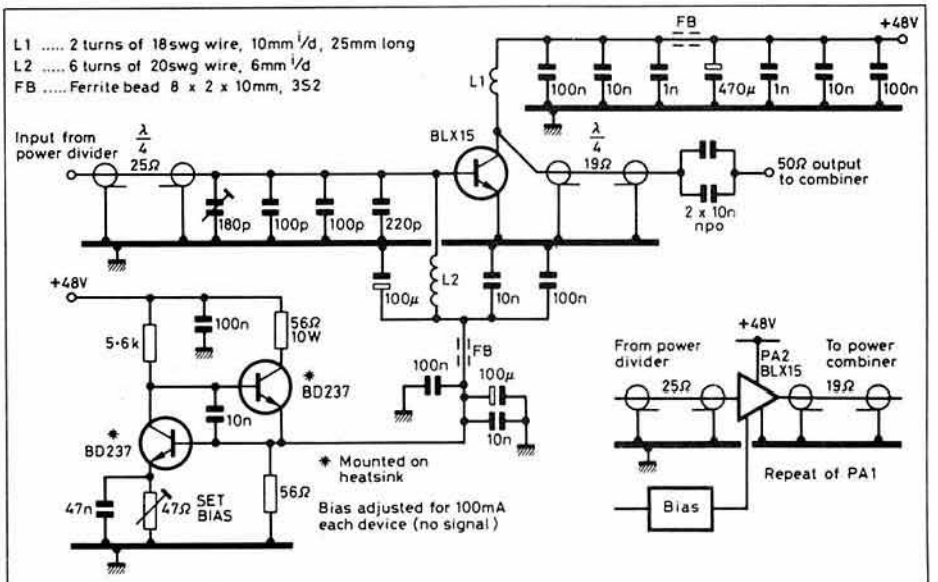


FIG 1B. POWER AMPLIFIER STAGE (PA1)

PA OUTPUT MATCH

Matching each individual output stage to 50ohms required by the power combiner is identical to methods used for input matching. The required impedance transformation is different, but the same principles apply. It is possible to obtain the required output load values from manufacturer's data, but for this type of output configuration it is better to depend on the impedance governed by collector voltage swing and power output requirement.

The output load, for one device, is defined by the following equation.

$$Rload = (Vcc - Vsat)^2/(2 \times Po) \quad \ldots 6$$

Where:
Rload is the required collector load (ohms)
Vcc is the applied collector supply voltage (V)
Vsat is the device saturation voltage (V)
Po is the required peak power output (W)

In order to design for good linearity it is important to ensure that collector voltage excursions do not approach the device saturation voltage. If this situation does arise through overdriving, or incorrect impedance matching at the output, severe distortion will occur. For this design, the value of Vsat will be assumed to be 6volts and calculations for output matching are based on 130W output from each stage. Supply voltage (Vcc) is 48volts, which allows a collector swing of 42volts.

From eq 6:

$Rload = (Vcc - Vsat)^2/(2 \times Po)$.

$\text{Therefore } Rload = (48{-}6)^2/(2 \times 130) = 6{\cdot}78\Omega.$

Now that the required collector load and final load impedance (50ohms) are known, it is possible to calculate the impedance required for a quarter wavelength transmission line transformer.

From eq 1:

$$Zline = \sqrt{50 \times 6{\cdot}78} = 18{\cdot}4\Omega \quad \ldots 7$$

Line length is calculated exactly as shown in eq 5. If ptfe cable is used dielectric constant and line length will remain the same.

The example calculation above does demonstrate one intrinsic poblem when quarter wave lines are chosen as the matching element. When matching two defined impedance using this method, non-standard cable impedances are often required. For the example output matching problem shown above, Belden 83307 ptfe screen cable was used which has an impedance of 19ohms. Unfortunately this particular cable is only available in 100foot lengths and at 50p/ft is prohibitively expensive when only a few feet are required.

All is not lost, however, for the required impedance may be achieved by paralleling lengths of standard miniature coax. For example, two lengths of UR110 50ohms coax gives 25ohms; three lengths gives 16·7ohms; two 50ohm cables and one 75ohm gives 18·75ohms. Either of these last two examples are close enough for the BLX15 design described.

Many solid dielectric screened cables are available, and although not intended for use as rf coaxial cable, do perform well if the insulating ▶

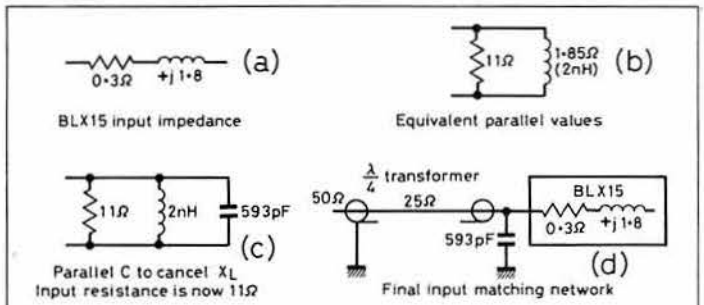
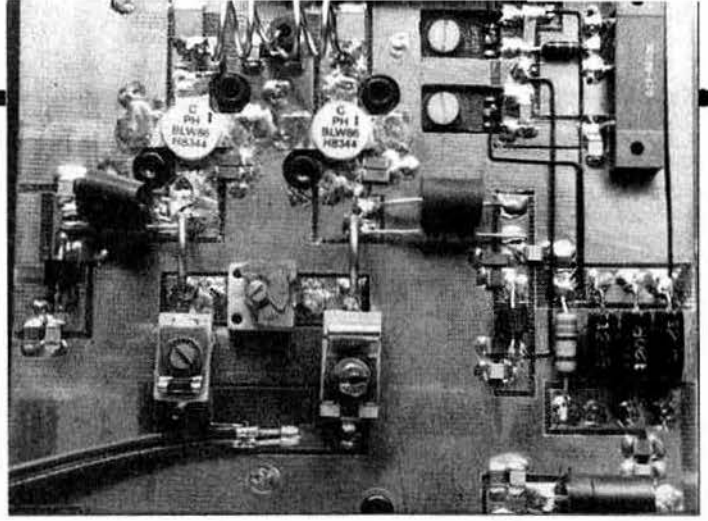


FIG 2. PA INPUT MATCHING ARRANGEMENTS



VIEW OF THE 75WATT DRIVER. BIAS COMPONENTS ARE IN THE TOP RIGHT-HAND CORNER. INPUT MATCHING AND COAXIAL TRANSFORMER ARE AT BOTTOM CENTRE.

dielectric properties are satisfactory. For example, many ptfe screened cables provide various 'odd' impedances combined with good rf performance. After finding unspecified screened cable, how can the approximate impedance be determined? With professional test equipment available the answer is quite simple: just measure it! Unfortunately these facilities are not available to most amateurs and another method must be found to give acceptable results.

The most simple method is to calculate Zo from physical cable dimensions and insulation dielectric constant (Er). All that is required is a micrometer to establish the inner and outer cable dimensions.

Zo may be calculated by using:-

$$Zo = 60/\sqrt{Er} \times 1\log 8\,(D/d) \quad \ldots 8$$

Where D is the inside diameter of the outer screen, and d is the outside diamter of the centre conductor. As an example, a length of ptfe screened cable was measured and found to have a characteristic Zo of 36ohms. Cable dimensions where then measured.

D = 0·043", d = 0·019". Dielectric constant of ptfe is 2·10 (polyethylene 2·28)

Using eq 8.

$$Zo = 60/\sqrt{2{\cdot}10} \times \log\,(0{\cdot}043/0{\cdot}019) = 33{\cdot}8\Omega.$$

This relates closely with the measured value. It provides a satisfactory and simple means of determining the characteristic impedance of unknown solid dielectric cables.

BASE BIAS

A low impedance bias supply is a very important feature which must not be overlooked on high power bi-polar linear amplifiers. Adequate current must be made available at all drive levels. Series feed resistors with diode clamp arrangements are frequently used for this purpose but this method should be used only on low power amplifiers.

The base bias circuitry used in this design is identical to that described in reference 1 (additional information is available in ref 3). It is temperature compensated, and provides adequate base current at all power levels. Inspection of the pa circuit diagram shows that individual, adjustable bias supplies are used on each output stage. There are two good reasons for this. The first is that each stage can be individually set for correct quiescent collector current. As the transistors in the output stage are not a matched pair, individual biasing allows each device to be set for optimum linearity. The other reason is that if a short circuit collector-base catastrophe should occur on one side, the remaining device is saved from a similar fate.

POWER DIVIDING AND COMBINING (HYBRIDS)

There are several established methods which may be employed to combine or divide power at 144MHz. Of these, Wilkinson hybrids, branch line and capacity coupled hybrids are the most frequently used. After considering the pros and cons of each type a decision was made to use a transmission line, capacitively coupled, 3dB hybrid. From a practical viewpoint, this type has several advantages.

(1) Unlike the Wilkinson, capacitively coupled hybrids use coaxial lines which are only 45° long. This makes a relatively compact combiner or divider which can be integrated onto the power amplifier board if rquired, using semi-rigid coax.

(2) With port impedances of 50ohms, coaxial lines used to make the hybrid are also 50ohms. If a Wilkinson type were chosen, 70·7ohm coax would be required with an electrical length of 90°. One disadvantage of capacitively coupled hybrids is the limited bandwidth of around 10%. However, for narrow band amplifier applications this can be advantageous in reducing even harmonics at the ouput.

The above discussion is not intended to denigrate the excellent properties of Wilkinson hybrids but serves to describe why an alternative was chosen for this particular application.

Apart from power dividing and combining functions, all types of hybrid provide isolation between ports. This feature is most important when combining the output (or inputs) of two independent, identical amplifiers.

Fig 4 shows a plot of isolation between the ports used to drive the inputs of the power amplifier stages (see Fig 1). From this it can be seen that isolation between bases of each output transistor will be around 50dB, so any mismatch at one transistor input will not affect the other. The same applies in the output combiner, isolation between collectors being assured.

Any imbalance of phase or amplitude between the two stages is 'dumped' into the 50ohm resistor placed at the isolation port. Should one half of the output stage fail, 50 per cent of the power from the remaining amplifier would be dissipated in the 50ohm isolation port resistor. The remaining 50 per cent is delivered to the combined output port as useful power. Of course, this statement is only valid if the failure of one amplifier has not demolished the power supply on the way out! This should not be the case in a correctly designed and protected psu, of course.

Port isolation shown in Fig 4 is the equivalent of input return loss. As return loss is directly related to vswr, it is possible to calculate the input vswr from the plot.

$$Vswr = 1 + \rho/(1 - \rho) \quad \ldots 9$$

$$\text{Return Loss} = 20\log_{10}\rho \quad \ldots 10$$

$$\text{Therefore } \rho = \text{Antilog}_{10}\,\text{Return Loss}/20 \quad \ldots 11$$

From Fig 4, the return loss at 145MHz is −54dB.

Therefore ρ = Antilog −54/20 =0·0019953

From eq 9. Vswr = 1+0·00199/1–0·00199 = 1·004:1

The plot shows graphically the relatively narrow bandwidth of capacitively coupled hybrids, although at 150MHz the return loss is still a satifactory 22dB. This equates to a vswr of 1·17:1.

All the above calculations assume 50ohm resistive terminations on each port.

Fig 5 shows the insertion loss at each output port of the power divider. This shows quite clearly the departure from 3dB for each port with changing frequency.

Those wishing to acquire more detailed information on various forms of hybrid couplers, well presented and informative literature may be found under References 4, 5, 6 and 7.

DIVIDER/COMBINER CONSTRUCTION

Construction of the input power divider and output combiner hybrids are identical and very simple to make. The coaxial lines are 45 electrical degrees long and made using 50ohms semi-rigid coax type UT141C (M17/130-00001).

Semi-rigid coax is preferred as it can conveniently be soldered directly onto copper laminate. Spacing of the two lines is not critical and is determined by the physical size of coupling capacitors used at each end of the lines. The value of coupling capacitors in divider and combiner is 22pF. For other values of coupling, different line lengths and coupling capacitor values are required (See Ref 7).

If good quality fixed value capacitors are not to hand for the ouput combiner, miniature air-spaced variables may be substituted. 30pF film dielectric trimmers are adequate for the input power divider. In all cases where variables are used, a means of setting the capacity to 22pF should be found before fitting. Alternatively, the hybrid may be optimized at 144MHz by using the following procedure. Refer to Fig 1 for

port designations and Fig 4 to see what should be achieved in terms of port isolation.

Terminate P4 with a 50ohm non-inductive resistor. P3 should already be terminated. Connect P2 to a 144MHz receiver via a 6dB attenuator. Apply a 50ohm signal generator to P1 and adjust the line coupling capacitors for minimum signal in the receiver. If a signal generator is not available, a strong local repeater may be used, but in order to ensure a 50ohm termination don't forget to place a 6dB pad in series with the antenna. Isolation optimization of the hybrid will be achieved using this simple method provided all ports are terminated with 50ohms. If semi-rigid coax is not readily available, ordinary 50ohm coax cable may be substituted. In all cases the electrical length is 45° for a 3dB capacitively coupled hybrid. Using the semi-rigid cable specified each line is 18cm long (velocity factor = 70%).

DRIVE AMPLIFIER

The 75W driver amplifier is of push-pull configuration using two BLW86 bipolar transistors operating from a 28V supply. The use of 28V transistors was dictated by what was available at rallies without spending vast sums of money! This has no disadvantages from an rf performance point of view at this power level, but does require a separate 5amp power supply. Input and output impedances are designed for 50ohms. Gain is approximately 10dB. Excellent linearity is obtained with 3rd order intermodulation products of −36dB wrt either tone at 70W pep output.

BIAS

Bias supply for the driver is similar to that used on the individual pa units, the major differences being that common bias is used for both halves of the push-pull pair. Quiescent current is set for a total of 160mA.

OUTPUT MATCHING

Matching for the output is based on a power output of 75W pep and a 28Vdc supply. It is the maximum power required to fully drive the power output stage. A quarter wave transmission line transformer is used to transform the collector to collector impedance up to 50ohms. This provides a reasonable match for the combining port of the pa power divider. In a push-pull amplifier the output impedance of each device is in series. This effectively multiplies the output impedance by a factor of 4 when compared to a single ended stage providing equivalent power output. Transformer step up ratios are thus smaller and matching made easier. This is particularly so with amplifiers operating from 12 or 24volt supplies.

Output load impedance for a push-pull amplifier is defined as:

$R\,load = (Vcc - Vsat)^2 / 0{\cdot}5 \times Pout$ . . .12

Where: Vcc = supply voltage. Vsat = saturation voltage.

Pout = Total output power.

As a collector voltage swing too close to the supply severely degrades linearity, Vsat should be limited to 6V.

From eq 12:

$Rload = (28-6)^2 / 0{\cdot}5 \times 75$. Therefore Rload = 12·9ohms

Referring to the text describing quarter wave coaxial line transformers and eq 1, it will be seen that a cable impedance of 25·3ohms is required to match 12·9ohms, to 50ohms.

This is obtained by paralleling 2 quarter wave lengths of 50ohm miniature coaxial cable (UR110).

This quarter wave section with one outer end grounded and the other end placed across the balanced collector loads provides the required 180° phase shift for correct push-pull operation. It also performs the function of a 'pseudo-balun', converting the balanced collector loads to 50ohms unbalanced. This simple piece of coaxial cable provides three functions – phase splitter, balun and impedance matching. The only other components required are a dc feed inductor, compensation capacitor (56pF) and dc isolation capacitors (10nF).

DRIVER INPUT MATCHING

Matching the input of the push-pull driver stage to 50ohms is accomplished in two separate stages. The first stage uses a 25ohm quarter wave transmission line transformer to reduce the 50ohm drive source impedance down to 12·5ohms. The principle is identical to that described earlier for output matching and in the same way, provides anti-phase signals to drive each base of the push-pull pair.

Additional components in the form of T networks are used to match the complex input impedance of each BLW 86 into 6·25ohms. The reason for this value being 6·25ohms is because the total impedance seen across the end of the coaxial transformer must be 12·5ohms. As the inputs of the push-pull stage are effectively in series, each half must be matched into 6·25ohms making a total for the pair of 12·5ohms.

The above statement, ie base to base impedance of the pair being twice that of the base emitter impedance of one device, only holds true for amplifiers operating between class A and AB. If the conduction angle is reduced to 180° or less; as in class C amplifiers, the base to ▶

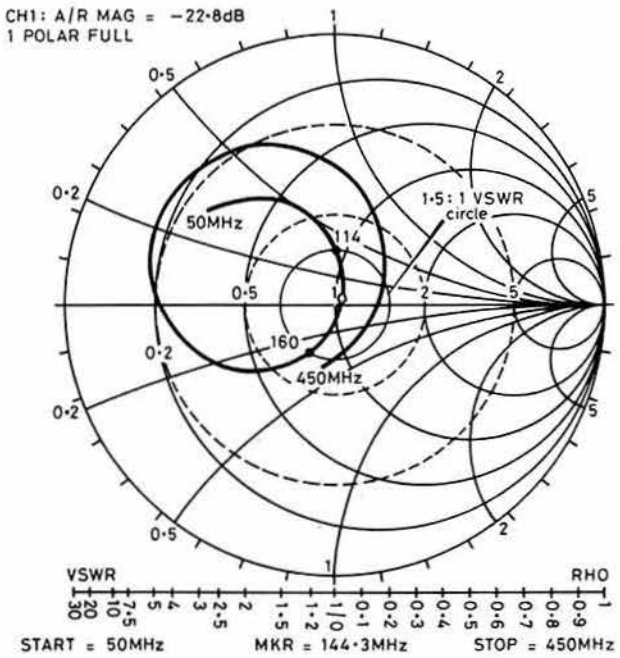


FIG 3. POLAR PLOT OF QUARTER-WAVE LINE TRANSFORMER CUT FOR 144MHZ

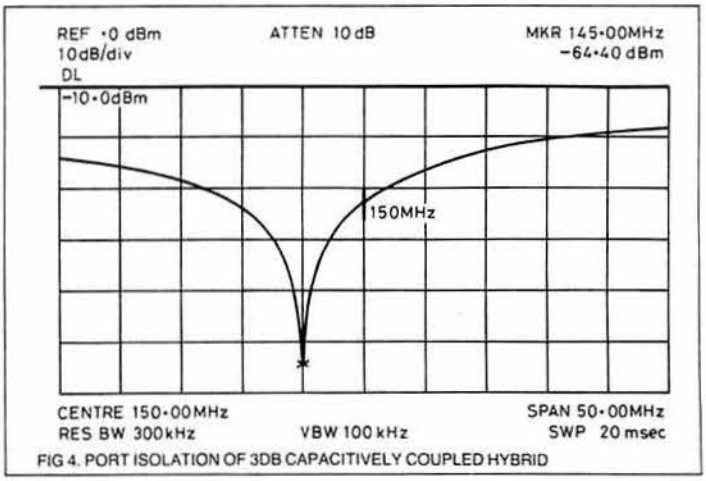


FIG 4. PORT ISOLATION OF 3DB CAPACITIVELY COUPLED HYBRID

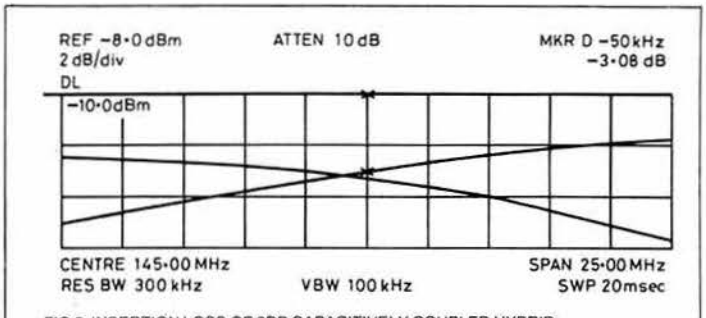


FIG 5. INSERTION LOSS OF 3DB CAPACITIVELY COUPLED HYBRID

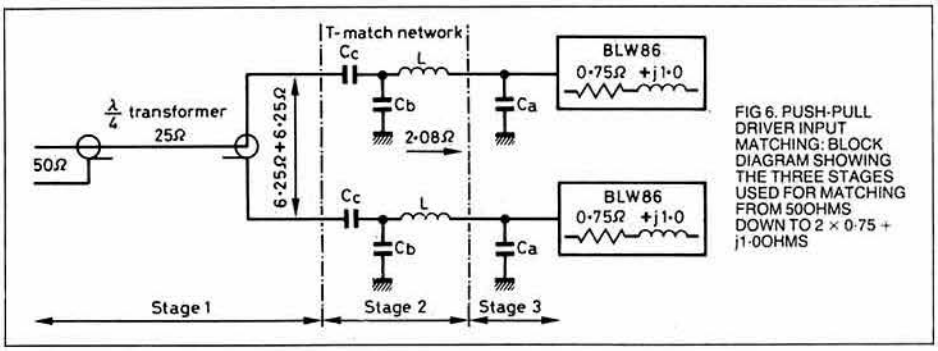


FIG 6. PUSH-PULL DRIVER INPUT MATCHING: BLOCK DIAGRAM SHOWING THE THREE STAGES USED FOR MATCHING FROM 50OHMS DOWN TO 2 × 0·75 + j1·0OHMS

base impedance will be approximately 4 times that of a single device.

In order to ease the remainder of input device matching, only one half at a time will be considered. When this is completed, values may be duplicated for the other half of the push-pull stage. A block diagram of the matching components required is shown in Fig 6.

Inspection of manufacturer's data for the BLW 86 shows a series component input impedance of 0·75 + j1·0ohms at 145MHz. It is a simple matter to cancel the inductive component (+j1·0) using a parallel capacitor at the base (Ca). The first step is to convert the series components of the input impedance into their parallel equivalent values.

The conversion formula and equivalent component values are shown diagramatically in Figs 8a and b. Calculated equivalent parallel values are shown in Fig 8b.

All that remains to complete stage 3 of the input matching is to cancel Xp (1·56ohms) with a capacitor of the same reactance. This capacitor (Ca Fig 7) resonates Xp at 145MHz, leaving a resistive input of 2·08ohms. Ca = 1 / 2 fXc therefore Ca = 703pF.

The final matching section shown in Fig 6 as stage 2, is solved using a simple 'LCC' T match network. This has to match 6·25ohms down to 2·08ohms and is identical for both sides of the push-pull driver amplifier. As the required transformation ratio is small (3:1) only one section is needed. L networks are used in preference to other methods as the Q in this type of network is controlled. The value of Q chosen is not particularly important, but must be low enough to give adequate bandwidth and practically realizable component values. For the example shown here a working Q of 4 was chosen. The method used for obtaining the required component values is fully explained in reference 1 and for brevity will not be repeated here.

Values for solving the final matching T network are:
R1 = 2·08ohms, R2 = 6·25ohms, QL = 4.
From these values: A = 2·158, B = 35·36.
After solving the equations based on these figures the final values for stage 2 (Fig 6) are.
Cc = 13·48ohms (81·4pF) Cb = 19·2ohms (57pF) and L = 8·32ohms (9·13nH).

In practice, capacitors Cb are placed in series and reduced to a single component. By so doing, the value is halved to 30pF. All capacitors in the T match are variable mica compression types based on the calculated values. This allows any variation of transistor input impedance to be adjusted out so presenting a low vswr to the 50ohm drive source. It was also found necessary to reduce the calculated values of Ca by 20% for optimum gain and match at 145MHz.

Although these procedures might seem rather daunting to the uninitiated, they are very straight forward and take far more time to describe than implement!

The driver amplifier described makes a simple, high performance power amplifier in its own right. At 75watts pep the devices are working well within their ratings and linearity is excellent (see Fig 7). Combiners and power splitters are eliminated and only a single regulated power supply capable of 5–6amps at 28V is required.

Like all rf power amplifiers an additional 6dB rf output requires 20dB more input in cost and construction effort!

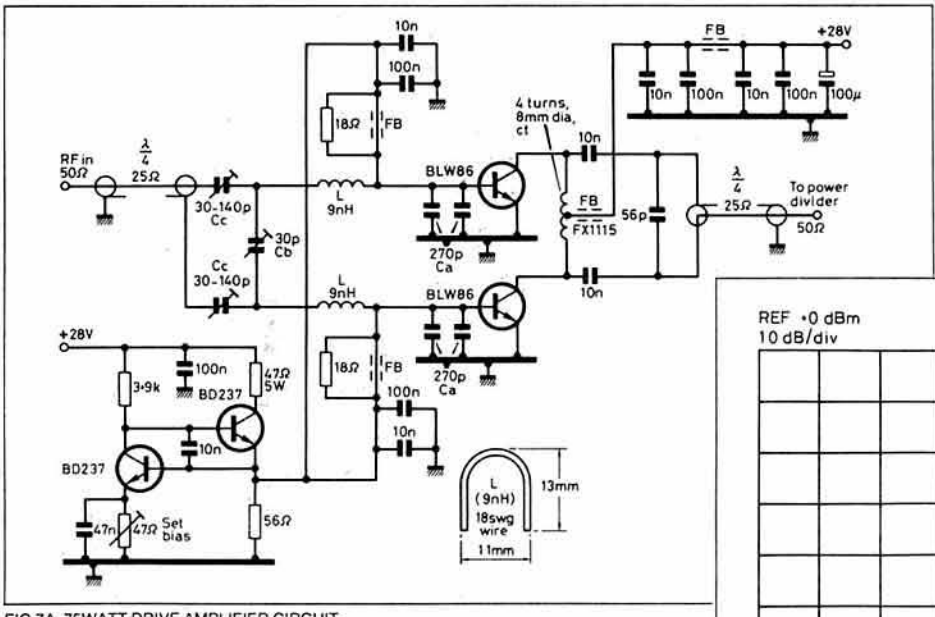
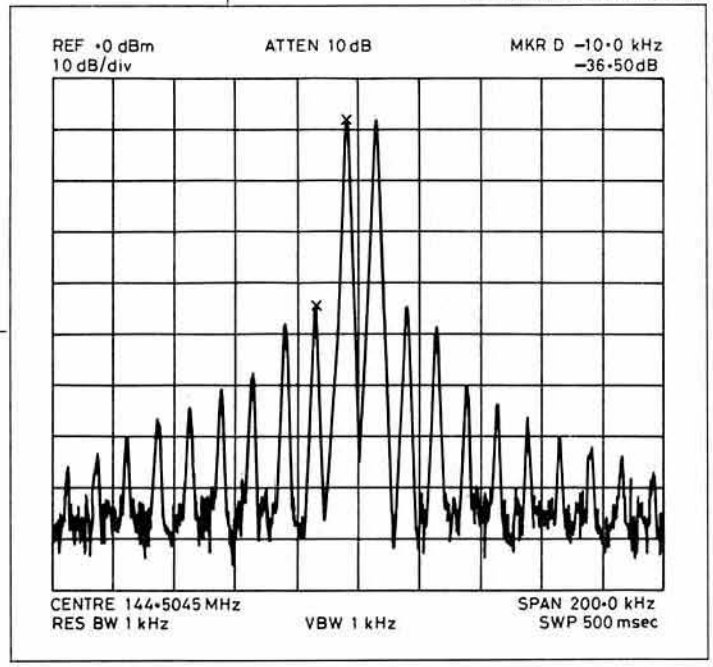


FIG 7A. 75WATT DRIVE AMPLIFIER CIRCUIT

TO BE CONCLUDED NEXT MONTH

FIG 7A NOTES
1) Ca, Cb, MICA COMPRESSION TYPES.
2) UNMARKED FB ARE 3S2 (BLUE).
3) COAXIAL TFMRS ARE 2 PARALLEL 35cm LENGTHS OF UR110.
4) 56pF COMPENSATION CAPACITOR IS GLASS OR ATC PORCELAIN TYPE.
6) ALL OTHER CAPACITORS, CHIP CERAMIC.
7) QUIESCENT COLLECTOR CURRENT SET FOR 160mA TOTAL.

FIG 7B. DRIVER AMPLIFIER IMD



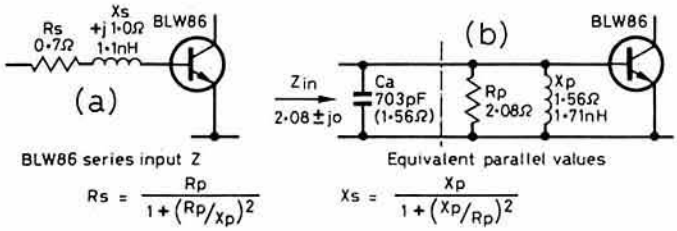
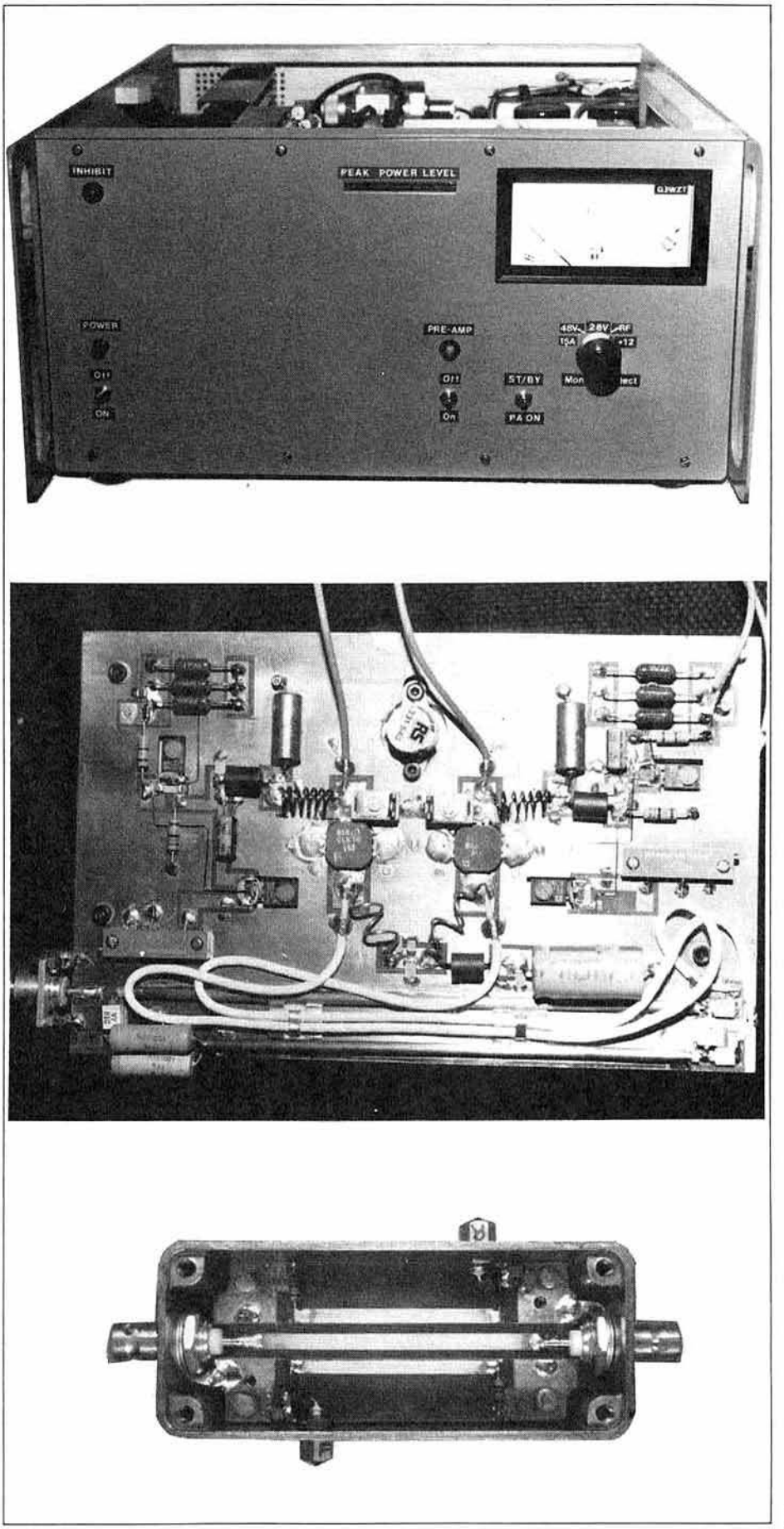


FIG 8. CONVERTING SERIES INPUT Z (a) INTO EQUIVALENT PARALLEL VALUES (b)

GUIDE-LINES

FOR THE DESIGN OF SEMICONDUCTOR VHF POWER AMPLIFIERS



PART TWO

CONTINUED FROM SEPTEMBER

BY

JOHN MATTHEWS

G3WZT

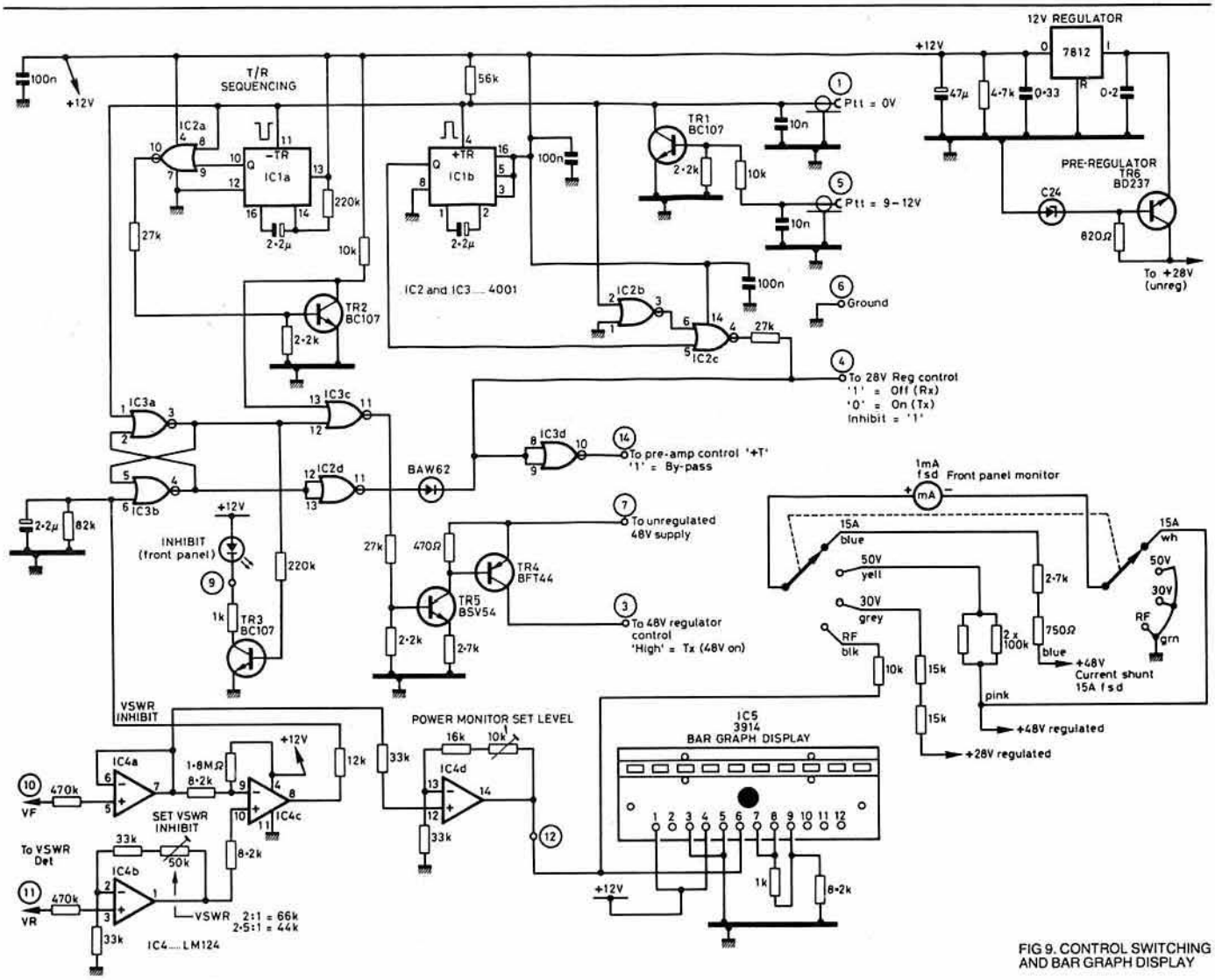


FIG 9. CONTROL SWITCHING AND BAR GRAPH DISPLAY

CONTROL AND SWITCHING

Fig 9 shows the circuit diagram for various functions required by the complete amplifier.

The functions included and the reasons for inclusion are:

1) SEQUENTIAL Tx/Rx SWITCHING

This function is controlled by mono-stables 9–IC1a, b. It ensures that when changing from Tx to Rx and vice versa, power supplies are switched in the correct sequence. All voltages are removed from the amplifier on receive by turning off the voltage regulators. The only exception is a low current 12V supply used for logic and the built-in Rx pre-amplifier.

When changing from Rx to Tx it ensures that antenna change-over relays (28V supplied) are energised before the main PA 48V supply. By so doing, rf is not present until the amplifier is correctly terminated by the antenna. The reverse is true when going from Tx to Rx and in both cases ensures that no rf is present when antenna change-over functions are taking place.

Two alternative inputs are available for ptt. One operates Tx functions when grounded, the other when a 6–12 volt positive supply is applied.

A correctly sequenced output to mute the receive pre-amplifier is available at 9–IC3 pin 10.

2) Tx INHIBIT

The purpose of Tx inhibit is to force the power supplies into Rx mode when high vswr is presented to the pa.

If for any reason a fault should develop in the antenna or feed system or, the amplifier is operated inadvertently without a load, full protection is provided for the output devices.

Forward and reverse voltages from the vswr detector (Fig 10) are buffered and amplified by 9–IC4a,b. For a preset value of vswr, differential amplifier 9–IC4c changes state and activates the inhibit latch 9–IC3a,b and 9–TR4 and 5 are switched off to remove the 48V supply. At the same time 9–IC2d output goes high, inhibiting the 28V regulator. A front panel LED driven by 9–TR3 is used to indicate vswr shutdown.

Pre-regulator transistor 9–TR6 is included to prevent input voltage to the 7812 12V regulator exceeding the recommended maximum when the 28V (pre-regulated supply) is lightly loaded.

3) BAR GRAPH DISPLAY

Although at first sight this might be thought of as a gimmick by some, it does represent the most accurate and effective method of displaying peak power. Moving coil meters will only indicate 30-50% of the true peak value at syllabic rates. This leads to the tendency of turning up rf drive to get readings which are comparable to steady carrier values.

Power level set potentiometer (10K) is set to illuminate nine sections on the bar graph display at 240W rf output. When the amplifier is driven with ssb, drive level is adjusted so that the final led (10th) is never illuminated. When correctly driven, this ensures that maximum peak power is never exceeded. Drive for the bar graph display is obtained from the forward voltage detector buffer 9–IC4a and amplifier 9–IC4d. The display is provided to indicate relative peak

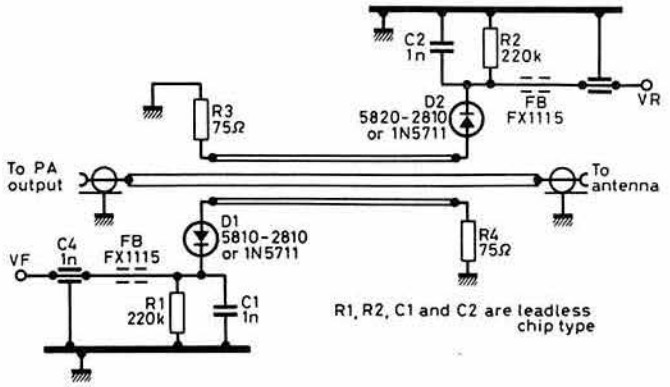


FIG 10. THE VSWR DETECTOR

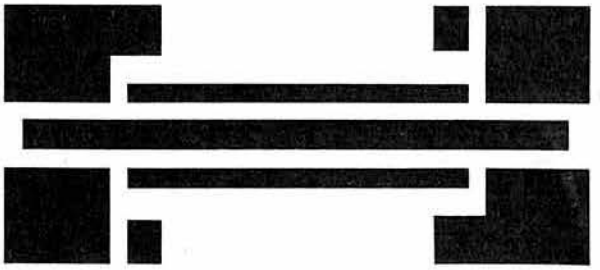


FIG 11. DESIGN FOR THE VSWR DETECTOR BOARD (TO SCALE)

power with a rapid response time. Any over driving on speech peaks can easily be seen and corrective action taken. Assuming the amplifier is operated into a well matched load the bar graph display may be calibrated in watts using a watt-meter as a calibration reference.

4) METERING

In addition to the bar graph display, a front panel meter was used to monitor the following parameters: 48V supply, 28V supply, 12V supply, relative power and pa supply current.

A current sense resistor in the 48V series regulator is used as a shunt for final collector current measurement.

VSWR BRIDGE

Details of the vswr bridge are shown in Fig 10 and the accompanying photograph. It is a simple directional coupler (voltage sample type) and provides two outputs, V forward and V reverse, which are fed to the operational amplifiers 9–IC4 and b.

If the decision has been made to inhibit the power amplifier when vswr exceeds 2:1, all that is required is a voltage ratio at the detector outputs for this particular value of vswr. The actual voltage values are not significant – it is the ratio between V forward and V reverse which must be known.

The ratio of VR/VF is a measurement of load reflection coefficient Rho,(ϱ) and is used to calculate vswr values.

The relationship between Rho(ϱ) and vswr is:
VR/VF = VSWR−1/VSWR+1
For a 2:1 vswr, VR/VF = 2−1/2+1. Therefore Rho = 0·333
Taking a nominal voltage for VF of 5V, VR will be 5 × 0·333 (1·66V) when VSWR is 2:1.

Now that relationships between VF, VR, and vswr are known, operation of the vswr inhibit circuitry may be more clearly understood. Referring to Fig 9, 9–IC4 is a voltage follower (buffer) stage and voltage at pin 7 is equal to VF.

If VF is 5V for a given power level, VR will be 1·66V at 2:1 VSWR. Amplifier 9–IC4b has a voltage gain of three making the dc level at pin 1 equal to that of pin 7 at 2:1 VSWR. 9–IC4c is a comparator. When the non-inverting input (pin 10) exceeds the level at pin 9, the output goes high, operating the inhibit latch which forces the amplifier into the receive mode. the inhibit function is reset by operating the ptt line.

Voltage for the bar display is obtained by amplifying the buffered voltage VF. Gain of 9–IC4d is variable and set to give 10V at full output power. Under these conditions, all 10 leds in the bar graph display are illuminated.

VSWR BRIDGE CONSTRUCTION

The vswr bridge is straightforward and easy to construct. A 50ohm transmission line and coupling loops are manufactured on 1·6mm double sided copper clad fibre glass board. Unwanted copper may be removed with a sharp knife and soldering iron or etched if preferred.

Board dimensions and etching details are shown in Fig 11.

Thes board is enclosed in a 90 × 38 × 32mm diecast box with connectors mounted at each end, which are soldered directly to the printed circuit transmission line.

DC outputs, VF and VR, are fed out via feedthrough capacitors to op-amp 9–IC4.

Width of the 50ohm transmission line is 3mm.

Coupling is approximately 30dB and the peak voltage at VF, into a well matched load, is approximately 5volts at 250W output.

POWER SUPPLIES

The 48V regulated supply is very similar to that described in reference 1. Current requirement is 12amps.

Foldback current limiting and simple over-voltage protection are provided. The mains transformer used was a 50V, 625VA toroidal type manufactured by ILP.

The 28V regulated supply is a conventional 723 arrangement driving a Darlington series pass transistor.

In order to protect expensive rf power transistors it is essential to include over-voltage protection and current limiting in both power supplies. If required, additional information on the power supplies used may be obtained from the author on receipt of an SAE.

RECEIVE PRE-AMPLIFIER

A receive pre-amplifier was included and may be switched out if not required. For this purpose a Wood and Douglas 144PA4/S was used. This particular pre-amp does possess a very original and effective diode switching system for power levels up to 30W. It is ideally suited as an add-on switchable pre-amplifier and may be easily incorporated into an existing pa.

A circuit diagram of the switching arrangement used in the Wood & Douglas 144PA4/S is shown in Fig 12. Reproduction is by permission of W&D.

The schematic diagram for Tx/Rx switching and Rx pre-amp is shown in Fig 13.

A three terminal 12volt regulator fed from the 28V pre-regulated line is used to supply the pre-amp. Although the 144PA4/S is rf switched, input T+ is also used to by-pass whilst in transmit.

CONCLUSION

As stated in the introduction the intention of this article was to present various guidelines and techniques used in vhf power amplifier design. It is not written as a constructional project.

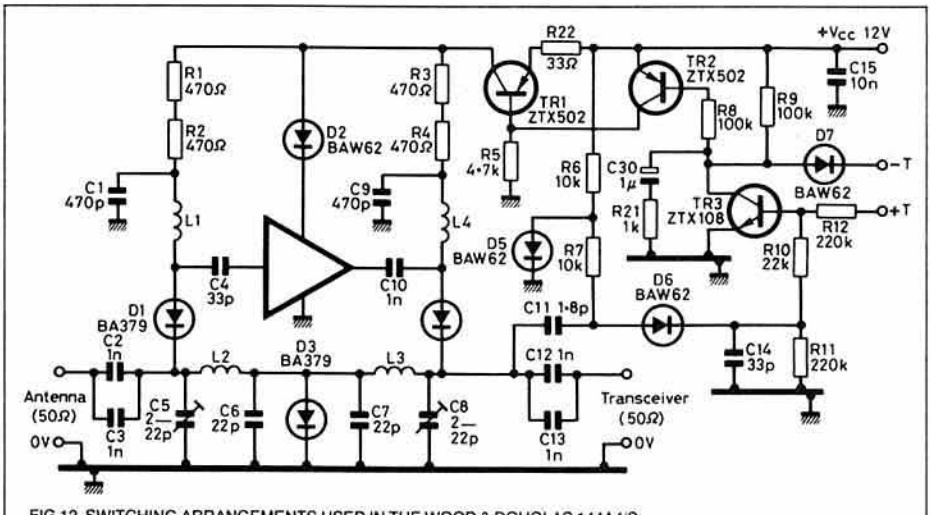
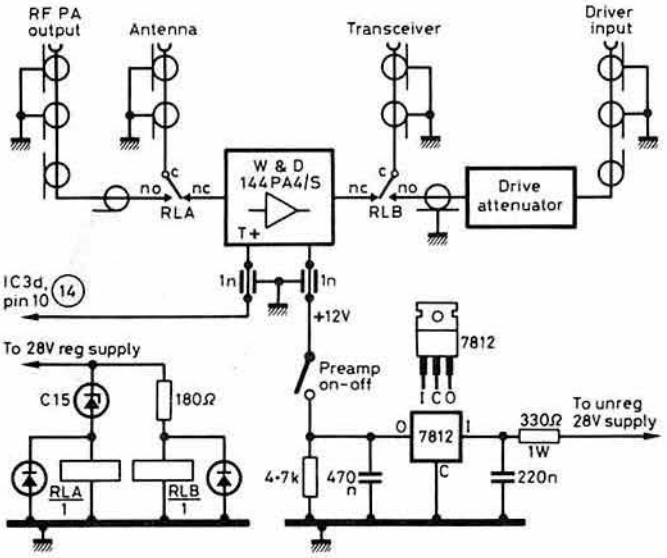


FIG 12. SWITCHING ARRANGEMENTS USED IN THE WOOD & DOUGLAS 144A4/S



▲ FIG 13. INPUT & OUTPUT SWITCHING PLUS PRE-AMP CONNECTIONS

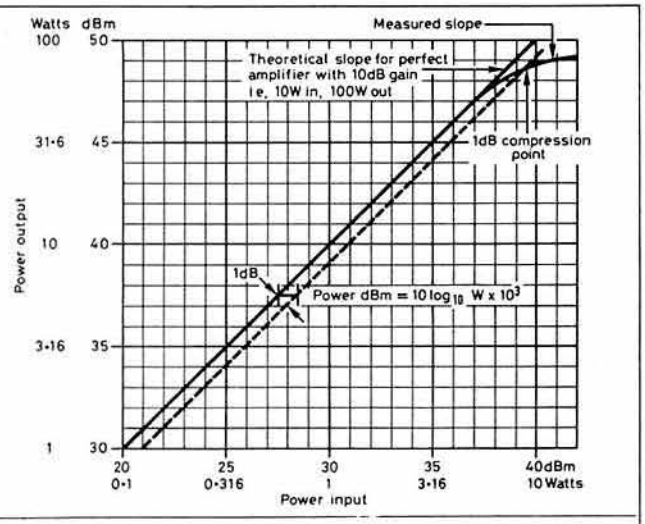

Watts dBm
Measured slope
Theoretical slope for perfect amplifier with 10dB gain i.e. 10W in, 100W out
1dB compression point
1dB
Power dBm = 10 log10 W x 10³
Power output
Power input
40dBm 10 Watts

FIG 14. EXAMPLE TO SHOW 1dB COMPRESSION POINT OF A LINEAR AMPLIFIER WITH 10dB GAIN ▲

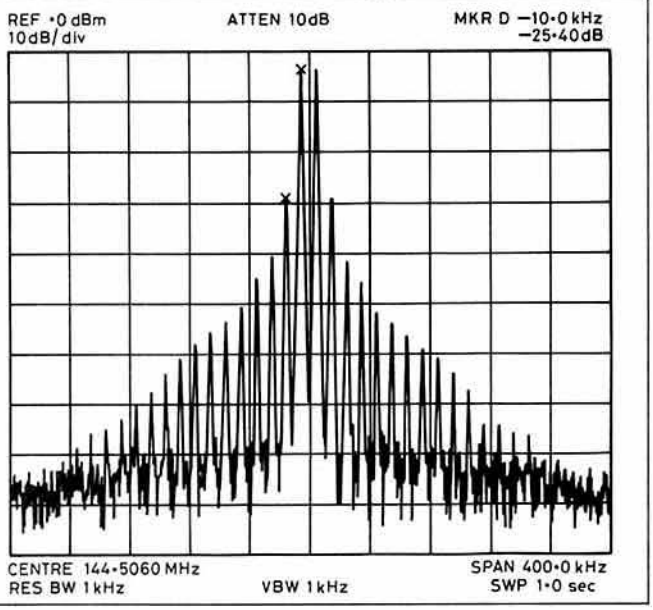


FIG 15. IMD PRODUCTS OF COMPLETED AMPLIFIER ▶

although sufficient information is included should experienced constructors wish to repeat any of the modules described. Like any design problems, whether it be analogue, rf, or digital, there is always more than one approach which may be taken to solve it. The various matching methods described here are no exception to this rule. They are, however, relatively simple to understand and implement.

In retrospect, there are some things which could be done to simplify this design. For example, two separate power supplies are used, one each for driver and power output stage. This could be simplified by using 48V devices all round.

Taking this argument one stage further, two modern generation high gain fets would provide 250-300W output and have sufficient gain to allow elimination of the driver stage. It quite simply depends on the depth of your pocket and the financial risks that you are prepared to take.

One 10μs flash with these devices works out at £15 per μs!

Solid state rf power amplifiers have unfortunately earned a poor reputation, particularly amongst devotees of valve amplifiers. In many cases this injustice is due to overdriving and attempting to get the advertised output power.

Get to know your pa by plotting a graph of input power against output power. Find out where the output starts to deviate from a straight line (compression) and don't apply more drive power than required to keep the amplifier operating in the linear region.

Once the amplifier performance is known, use a bar graph display to ensure that maximum linear power levels are never exceeded.

The graph shown in Fig 14 is hypothetical only and serves to demonstrate a simple method of characterising an amplifier using two power meters and a dummy load.

Power is shown plotted in dBm and serves as a convenient means of demonstrating the 1dB compression point.

The amplifier described is capable of delivering 250W, but at this level linearity is at the limit of acceptable performance with 3rd order imd products of −20dBc wrt either tone. A plot of 2 tone performance at 200W pep output is shown in Fig 15. At this level 3rd order imd products are −25dBc wrt either tone and represents a very acceptable level of linearity.

As a comparison, a commercially manufactured 10W in, 180W out, 144MHz pa was tested. At 125W pep the 3rd order imd products were −24dBc and at 150W, −15dBc. This is not a very impressive result, and all the more reason to "get to know your pa" and keep it clean by driving at the correct power level. From a constructional point of view, the described pa does represent a fairly advanced project.

It was never intended to compete with the straightforward and well proven approach of a 4CX250 amplifier. Hopefully what it does is provide an alternative approach and describes some methods which may be employed in the design of solid state vhf linear amplifiers.

ACKNOWLEDGEMENTS

I would like to than MEL for giving permission to use their test equipment.

Also my thanks go to Tony Ault G3KTU and Tony Wadsworth G3NPF for reviewing the text and providing constructive criticism during preparation.

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